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RESUME

La détection des signaux à large bande est difficile dans la technique sonar, à cause des incertitudes de la puissance de bruit à large bande et, par conséquent, de l'incertitude quant au seuil de détection nécessaire. Toutefois, si le récepteur est suffisamment directif, il est possible d'exploiter les caractéristiques spatiales du champ à large bande pour isoler les signaux d'onde plane du bruit de fond plus diffus. Le présent document étudie la mise en paramètres de l'estimation du gisement à large bande provenant d'un récepteur de réseau linéaire, en modes complexes (méthode de Prony). L'identification des modes associés aux ondes planes d'arrivée est examinée comme un moyen de séparer directement les signaux du bruit. On suppose être en présence de faisceaux à large bande seulement, de sorte que les estimations de gisement sont influencées par la réponse du réseau. Les questions du nombre de modes et des erreurs d'estimation sont traitées au moyen d'exemples.

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# NOISE NORMALIZATION OF BROADBAND SONAR DATA IN BEARING SPACE

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## Abstract

Detection of broadband signals is difficult in the sonar application due to the uncertainty in the broadband noise power and hence the uncertainty in the threshold required for detection. However, if the receiver has sufficient directivity, then it is possible to exploit the spatial characteristics of the broadband field to isolate plane wave signals from the more diffuse noise background. This paper considers a parameterization of the broadband bearing estimate from a line array receiver in terms of complex modes (Prony method). Identification of the modes associated with the plane wave arrivals is examined as a means of directly separating the signals from the noise. Only broadband beampowers are assumed available, so the bearing estimates are contaminated by the array response. Issues of mode number and estimation error are addressed by way of example.

## I. Introduction

Broadband detection is increasingly attractive in sonar. Because of the large time-bandwidth products available, signals having relatively small signal-to-noise ratios are theoretically detectable. However, practical detectors seldom approach theoretical performance because the spectral and spatial character of the noise is often unknown. Noise normalization refers to the process of estimating the background noise power, thereby allowing a constant threshold to be used for signal detection.

The problem of normalization in bearing space for broadband plane wave detection is similar to normalization for narrowband detection in the frequency domain. The latter typically normalizes against the average noise power in adjacent frequency cells. Such averaging is often unsuitable in bearing space because of the small number of independent beams available and the non-whiteness of the noise across beams. However, when a spatial array of reasonably high directivity is used, then it is possible that the spatial structure of the broadband field can be exploited in the normalization process.

In this paper, we consider normalization in bearing space for a towed line array receiver. A modal decomposition (related to the Prony method[1],[2]) of the Broadband Bearing Display(BBD) is examined as a means of directly identifying the plane wave components. Parameters yielded by the decomposition are mode angles, radii and weights. Since it is desirable that normalization occur at the post-processing level, only the beampower estimates are assumed available. As a result, field directionality will be corrupted by the array response, and plane wave signals will no longer appear as discrete components. Note that by directly identifying the plane wave signals, the method is in fact a detection device.

The procedure essentially treats the BBD as the positive-frequency spectrum of a band-limited process, from which an autocorrelation function (estimate) is obtainable. Hence, recent developments[3] in autoregressive-moving average(ARMA) modelling that exploit a singular value decomposition of the extended-order autocorrelation matrix can be used for identification of the modes.

## II. The Broadband Bearing Display

A schematic of the classical signal processing used to produce the Broadband Bearing Display(BBD) is shown in Figure 1. Narrowband beams are formed by Fourier analysis of the line array sensor signals, followed by phase-to-a-plane beamforming of the Fourier coefficients in each frequency cell over the band of interest. The resultant beampowers for each beam angle are integrated over time, T, then summed over frequency band, B, where the powers can be weighted in a manner to optimize some detection criterion (e.g. Eckart filter). The bandwidth-time product, BT, reflects the stability of the beampower estimates.

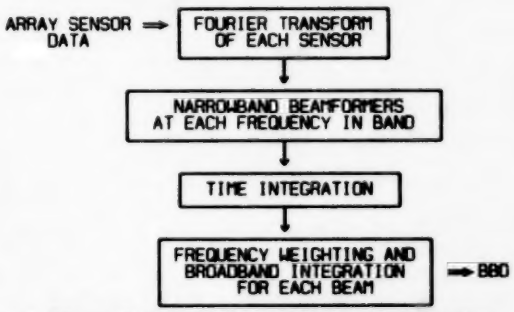
ARRAY SENSOR DATA
FOURIER TRANSFORM OF EACH SENSOR
NARROWBAND BEAMFORMERS AT EACH FREQUENCY IN BAND
TIME INTEGRATION
FREQUENCY WEIGHTING AND BROADBAND INTEGRATION FOR EACH BEAM
BBD

**Figure 1. Schematic of BBD signal processing.**

Figure 2 contains an example of the ideal BBD ($BT\to\infty$) for a line array of 48 uniformly spaced sensors in a computer generated broadband acoustic field spanning the octave band up to the frequency at which the inter-sensor separation is 1/2-wavelength (aliasing frequency). The abscissa, $\beta$, is the sine of the steering angle relative to array broadside; note that it is possible to measure energy in *virtual beams*, ($|\beta|>1$), resulting, for example, from leakage or spatially uncorrelated noise. In this case, the noise field consists of spherically and cylindrically isotropic noise, as well as uncorrelated noise, in proportions representative of the undersea acoustic field. Five temporally-white plane wave signals have been inserted, with arrival angle, $\beta$, power, P, and beam signal-to-noise ratio (S/N) tabulated for each signal beside the figure. The strong signal at $\beta$=0.985 is the energy received from the tow ship.



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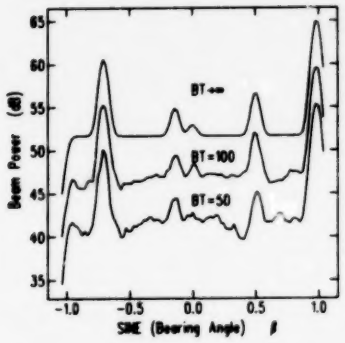


| β | P(dB) | S/N(dB) |
|---|---|---|
| -.707 | 60 | 8 |
| -.139 | 52 | 0 |
| .000 | 47 | -5 |
| .500 | 55 | 3 |
| .985 | 65 | 13 |

**Figure 2. Examples of simulated BBD's.**

Figure 2 also contains two realizations of BBD's obtained for finite BT, namely BT=100 and BT=50, and offset successively by -5 dB for clarity. For finite BT and Gaussian statistics, the beampower estimates will exhibit a chi-square density with of the order of 2BT degrees of freedom. Hence, we have assumed the mean beampower to equal $S(\beta)$, the power level in the ideal case, with a variance given by $S^2(\beta)/BT$. For moderately large BT, a Gaussian density having this mean and variance is a valid approximation to the chi-square density. For the purposes of this paper, the interbeam correlation introduced by the beamformer is modelled as a simple exponential function. Hence the spectra of Figure 2 have been generated by adding to the mean power level $S(\beta)$ a random Gaussian vector with correlation matrix elements:

$$E[\hat{S}(\beta_n)\hat{S}(\beta_m)] = \frac{S(\beta_n)S(\beta_m)}{BT}\,\rho^{|n-m|}. \qquad (1)$$

A correlation coefficient of $\rho$=0.8 was chosen for the data of Figure 2 so that significant correlation occurs for beam spacings less than about 5.

### III. The Modal Decomposition

We consider the broadband bearing response, $S(\beta)$, as the positive frequency spectrum of a band-limited stationary process, defined on the band $-2\beta_M \le \beta \le 2\beta_M$, $\beta_M \ge 1$. Note that $\beta_M$ may exceed unity in order to contain energy leaked into the virtual domain.

Given the above equivalence, we define the discrete autocorrelation sequence, $r_n$, as the inverse discrete Fourier transform of the sampled BBD. In order to avoid aliasing in the autocorrelation domain the number of beams computed in $S(\beta)$ must be sufficiently large. Consideration of the narrowest possible feature in $S(\beta)$ leads to the constraint that the number of beams must exceed twice the number of elements in the array. The sequence, $r_n$, may be approximated by a sum of complex modes, where the pth-order modal approximant, $r_n^{(p)}$, is[2]:

$$r_n^{(p)} = \sum_{m=1}^{p} A_m z_m^{|n|}, \quad n=0,\pm1,\pm2\ldots., \qquad (2)$$

$$z_m = \alpha_m e^{j\pi\beta_m/2\beta_M},$$

and the sampling frequency is $4\beta_M$. The $z_m$ is the mth complex mode with arrival angle, $\beta_m$, radius, $\alpha_m$, and (complex) mode weight $A_m$. With $r_n^{(p)}$ real, the modes and weights are either purely real, or appear in complex conjugate pairs (note that $r_n^{(p)}$ is real here due to the symmetry imposed on $S(\beta)$, which is not explicity necessary).

Since $S(\beta)$ is continuous due to the effect of the array response, all radii are bounded by $\alpha_m < 1$, and the modal spectrum $S(\beta)^{(p)}$ may be written:

$$S(\beta)^{(p)} = \sum_{m=1}^{p} A_m \frac{1-z_m^2}{(1-z_m e^{-j\pi\beta/2\beta_M})(1-z_m e^{j\pi\beta/2\beta_M})}. \qquad (3)$$

$S(\beta)^{(p)}$ will be real and symmetric; however, it is not necessarily nonnegative since the mode weights are not purely positive real.

The spectrum of a pair of conjugate modes exhibits the typical damped second order response and thus appears a reasonable choice as a basis function for the identification of the plane wave arrivals. It is reasonable to expect that modes having $\alpha_m$ near unity would model the plane wave components, whereas modes with $\alpha_m$ near zero would model the more diffuse background.

### IV. Modelling Procedure

The modelling procedure is related to the Prony method [1], which fits a sum of complex exponentials to a finite sequence. In this case, the sequence is the autocorrelation sequence, $\{r_0, r_1 \ldots\}$, with at least 2p terms available. The essence of the procedure stems from the realization that the $z_m$ for the p-th order fit to the given sequence may be obtained as roots of the polynomial:

$$a_0 + a_1 z^{-1} + \ldots\ldots + a_p z^{-p}, \qquad a_0 = 1 \qquad (4)$$

where the (p+1)x1 vector, a, obeys the set of difference equations:

$$\sum_{k=0}^{p} a_k r_{n-k} = 0, \qquad n \ge p. \qquad (5)$$

Once the $z_m$ are found, the weights, $A_m$, can be obtained by solving the linear system expressed by (2), with the actual data $r_n$ in place of $r_n^{(p)}$. While it is possible to constrain the weights to be positive real, the resultant spectral fit to $S(\beta)$ will generally be poorer than for unconstrained $A_m$.

Eqn. (5) is recognized as a subset of the *extended Yule-Walker* equations which characterize an ARMA(p,q) model (causal) for q<p, where a is the vector of p+1 autoregressive parameters ($a_0$=1). Hence, we may call upon recent developments[3] in ARMA modelling to estimate the "best" order, p, and the corresponding a. Of course, given exact knowledge of the $r_n$ and of the correct p, only p equations (for example, (5) with $p \le n \le 2p$ ) need be solved to obtain a. However, when the $r_n$ contain measurement error, it has been found beneficial to work with an overdetermined system of t equations, t>p (the terminology below closely follows [3]):

$$\mathbf{R}\,\mathbf{a} = \boldsymbol{\Theta}. \qquad (6)$$

$\boldsymbol{\Theta}$ is a tx1 zero vector and R is the tx(p+1) overdetermined ARMA autocorrelation matrix. With $r_n$ available from $r_0$ to

$r_{p+1-i}$, R defines a set of contiguous equations when its terms have the form $R_{mn} = r_{p+n-m}$. If R is full rank (*i.e.*p+1), then the underlying process has order greater than p, and there is no non-trivial solution to (6) (measurement error may also make R full rank, although the process has order less than p). However, a "best fit" a can be obtained in the least-squares sense by solving:

$$R^* R\, a = \gamma\, e, \qquad (7)$$

where e is a (p+1)x1 vector with first term unity and zero otherwise, and $\gamma$ is selected so that $a_0$=1.

Of more interest in this application is the case where R is less than full rank (or has an *effective* rank less than full) since one then has a measure of assurance that the modal decomposition has a sufficient number of modes to model the underlying process. Thus we will attempt to select p sufficiently large so that R is less than full rank and turn to the singular value decomposition(SVD) of R as a means of estimating the number of modes required. The SVD of R is given by:

$$R = U \Sigma V^*, \qquad (8)$$

where U and V are txt and (p+1)x(p+1) unitary matrices whose columns contain the eigenvectors of $RR^*$ and $R^*R$, respectively. $\Sigma$ is a tx(p+1) diagonal matrix where $\Sigma_{kk}$, k=1,p+1 are the singular values of R arranged in order of decreasing magnitude, which in turn are the positive square roots of the eigenvalues of $R^*R$. If R has rank $p_0<p+1$, there will be $p-p_0+1$ zero singular values. If, on the other hand, there are $p-p_0+1$ singular values *approximately* zero the effective rank of R can be assumed to be $p_0$. Note that the singular values may be near zero due to measurement error for the order $p_0$ process, or because the order $p_0$ process is a sufficently accurate approximation to the true process. We present examples of rank selection in the next section.

Given $p_0$, the rank $p_0$ optimum approximation to R is:

$$R^{(p_0)} = U \Sigma^{(p_0)} V^*, \qquad (9)$$

where $\Sigma^{(p_0)}$ is produced by setting the smallest $p-p_0+1$ singular values of $\Sigma$ to zero.

Having obtained a reduced rank correlation matrix, $R^{(p_0)}$, an autoregressive parameter vector of length $p_0+1$ should exist. In principle, $p-p_0+1$ such solutions can be obtained directly from (6) by replacing R with $R^{(p_0)}$ and utilizing the dimensionality of the null space of $R^{(p_0)}$, to constrain the solution in an appropriate manner. Alternatively, if we define $R_k^{(p_0)}$ as the submatrix of $R^{(p_0)}$ consisting of the $k$th through $(p_0+k)$th columns then:

$$R_k^{(p_0)}\, a = \Theta, \quad 1 \le k \le p-p_0+1, \qquad (10)$$

yields theoretically the same solution for a of reduced dimension $(p_0+1)$, for each of the $p-p_0+1$ possible values of k. However, due to computational noise in the formation of $R^{(p_0)}$, a unique solution for all k will generally not exist. The best least squares solution is the solution to:

$$S^{(p_0)}\, a = \gamma\, e, \qquad (11)$$

where $$S^{(p_0)} = \sum_{k=1}^{p-p_0+1} R_k^{(p_0)*} R_k^{(p_0)},$$

and $\gamma$ and e are defined as in (7). Once a is obtained, the modes result from the rooting of (4), and the mode weights from solution of the linear system, (2), where $r_n^{(p)}$ is replaced by the appropriate entries of $R^{(p_0)}$.

## V. Examples of the Procedure

We now demonstrate the identification procedure using three sets of examples, one for each BBD in Figure 2. These BBD's result for a 48 sensor line array from which we simulate 103 beams over one octave in frequency. The limits on the input broadband spectrum are defined by $\beta_M$=1.037. In all cases we use an initial order, p=29, and 70 (t=70) extended Yule-Walker equations in the analysis. A least squares solution was obtained for the autoregressive parameter vector via (11). The singular values for each BBD are presented in Table 1, assuming R is normalized so that $r_0$=1. The values decrease in a smooth, monotonic fashion, with a clear choice for $p_0$ not apparent. As BT decreases and hence the measurement noise increases, the small singular values become larger, further complicating the selection of $p_0$.

| | | | | | |
|---|---|---|---|---|---|
| BT=∞ | | | | | |
| 3.00811 | 0.33925 | 0.08159 | 0.02925 | 0.00360 | 0.00080 |
| 1.99712 | 0.22355 | 0.07962 | 0.02106 | 0.00318 | 0.00034 |
| 1.01703 | 0.19810 | 0.07145 | 0.01597 | 0.00273 | 0.00027 |
| 0.87286 | 0.13561 | 0.04063 | 0.00691 | 0.00267 | 0.00020 |
| 0.39668 | 0.09407 | 0.03519 | 0.00628 | 0.00099 | 0.00009 |
| BT=100 | | | | | |
| 2.87104 | 0.37917 | 0.08478 | 0.04054 | 0.01422 | 0.01079 |
| 1.94541 | 0.26094 | 0.08395 | 0.03043 | 0.01384 | 0.00900 |
| 1.00013 | 0.20395 | 0.07365 | 0.02837 | 0.01363 | 0.00903 |
| 0.86312 | 0.12184 | 0.06067 | 0.02155 | 0.01303 | 0.00717 |
| 0.42611 | 0.09798 | 0.05067 | 0.01952 | 0.01137 | 0.00668 |
| BT=50 | | | | | |
| 3.39015 | 0.33526 | 0.08677 | 0.03211 | 0.02734 | 0.01788 |
| 2.26714 | 0.16698 | 0.08513 | 0.03147 | 0.02629 | 0.01398 |
| 0.89324 | 0.16552 | 0.07525 | 0.03055 | 0.02379 | 0.01208 |
| 0.77010 | 0.12792 | 0.05752 | 0.03005 | 0.02094 | 0.01073 |
| 0.39153 | 0.09461 | 0.04734 | 0.02880 | 0.01816 | 0.01036 |

**Table 1. Singular Values for BBD's in Figure 2, p=29.**

Figure 3 and Table 2 contain the results of the modal decomposition for each BT selection. In each case, $p_0$ has been set at 8, 15 and 20. The table summarizes the mode angles and radii identified for each $p_0$, and the magnitude, P(in dB), of the individual equivalent spectral component for the mode (plus its conjugate pair) evaluated at the mode angle. A minus sign has been used to indicate those modes which have negative spectrum values, as is possible when the mode weight, $A_m$, is not constrained to be positive real. Those modes not occurring at $|\beta|=\beta_M$=1.037 have a conjugate pair that is not tabulated. The figure shows $S(\beta)^{(p_0)}$ as expressed by (3) for the three $p_0$ values and particular BT. In addition, the locations of the mode angles whose radii exceed $\alpha_m$>0.5 are indicated below each reconstructed spectrum. The direction of the marks, up or down, indicates whether the mode spectrum at the mode angle is positive or negative. The constraint on $\alpha_m$ removes those modes that model the diffuse background as opposed to the plane wave arrivals.

Generally, these results reveal that if too low a $p_0$ is chosen from the SVD, not enough modes are available to identify all the plane wave arrivals. In this example, we would expect to need at least 5 pairs of modes ($p_0$=10) close to the unit circle to model the signals, while a few modes well removed from the unit circle would be necessary to model the background. As $p_0$ increases, the strongest signals are the first to be identified. For example, as shown

in Figure 3a for BT-∞ and $p_0$=8, the four mode pairs obtained accurately estimate the angles and powers of the four strongest plane waves, missing the weak signal at $\beta$=0.000. As $p_0$ increases further, more modes are available to model $r_{nn}$. One would hope that these new modes would be used to identify signals which had remained undetected at a lower $p_0$. However, since a single mode does not exactly model a plane wave arrival, these additional modes are first used to refine the approximation in the areas of large power content. Eventually, increasing $p_0$ leads to identification of the low level signals. For example, it is necessary to increase $p_0$ to 15 in Figure 3a before the weakest signal at $\beta$=0.000 is identified. However, this also produces a negative magnitude mode at $\beta$=-0.693, as well as others at the extremities of the band to model the background.

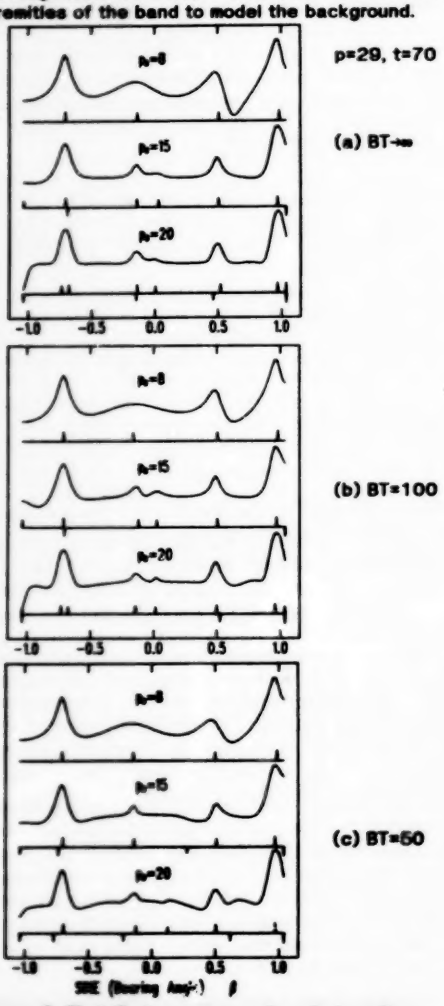


**Figure 3. Modelled spectra and mode locations.**

If $p_0$ is increased again, extraneous modes are introduced (see Figure 3a for $p_0$=20). A common result of excessive $p_0$ is the splitting of modes identifying plane wave arrivals into positive-negative or positive-positive pairs. While the combined effect of these pairs improves the spectral approximation, a straightforward interpretation of the plane wave arrivals can no longer be obtained directly from the mode parameter list. In such a situation, an indirect interpretation in terms, for example, of the gradient of $S(\beta)^{(p_0)}$ is a possible, albeit more difficult, alternative.

| | BT-∞ | | | BT=100 | | | BT=50 | | |
|---|---|---|---|---|---|---|---|---|---|
| | $\beta_m$ | $\alpha_m$ | P(dB) | $\beta_m$ | $\alpha_m$ | P(dB) | $\beta_m$ | $\alpha_m$ | P(dB) |
| $p_0$=8 | | | | | | | | | |
| | -0.700 | .95 | 60.9 | -0.711 | .95 | 60.6 | -0.707 | .95 | 60.2 |
| | -0.140 | .77 | 53.3 | -0.169 | .66 | 52.9 | -0.151 | .71 | 53.0 |
| | 0.498 | .91 | 55.5 | 0.491 | .93 | 56.1 | 0.485 | .89 | 53.2 |
| | 0.970 | .95 | 65.1 | 0.966 | .95 | 64.8 | 0.968 | .95 | 65.5 |
| $p_0$=15 | | | | | | | | | |
| | -1.037 | .91 | 39.6 | -1.037 | .88 | 47.8 | -1.037 | .88 | 53.2 |
| | -0.708 | .94 | 60.9 | -0.711 | .94 | 60.6 | -1.037 | .90 | -35.5 |
| | -0.693 | .66 | -54.0 | -0.706 | .79 | -52.9 | -0.738 | .74 | -53.6 |
| | -0.144 | .95 | 52.1 | -0.126 | .94 | 51.2 | -0.705 | .95 | 60.1 |
| | 0.023 | .91 | 46.3 | 0.016 | .91 | 49.0 | -0.147 | .97 | 51.2 |
| | 0.492 | .95 | 55.2 | 0.491 | .95 | 55.6 | 0.276 | .83 | -39.9 |
| | 0.956 | .95 | 64.4 | 0.953 | .95 | 64.2 | 0.497 | .95 | 52.8 |
| | 1.037 | .87 | 60.5 | 1.037 | .96 | 51.8 | 0.957 | .94 | 64.7 |
| | 1.037 | .66 | 51.8 | 1.037 | .96 | -57.8 | 1.037 | .95 | -59.2 |
| $p_0$=20 | | | | | | | | | |
| | -1.037 | .18 | 52.6 | -1.037 | .93 | -51.9 | -1.037 | .96 | -46.8 |
| | -1.037 | .93 | -51.4 | -1.037 | .75 | 50.8 | -0.776 | .93 | -52.2 |
| | -0.736 | .93 | 51.8 | -0.733 | .94 | 56.4 | -0.698 | .95 | 59.8 |
| | -0.682 | .93 | 54.5 | -0.676 | .93 | 49.0 | -0.249 | .43 | 53.6 |
| | -0.151 | .90 | -49.4 | -0.149 | .88 | 47.2 | -0.231 | .86 | -48.2 |
| | -0.145 | .93 | 53.6 | -0.145 | .95 | 50.3 | -0.131 | .95 | 50.0 |
| | -0.002 | .96 | 45.4 | 0.007 | .98 | 46.4 | 0.116 | .95 | 38.2 |
| | 0.452 | .92 | -49.3 | 0.487 | .95 | 56.4 | 0.490 | .93 | 54.3 |
| | 0.505 | .94 | 54.5 | 0.509 | .86 | -51.7 | 0.806 | .87 | -48.3 |
| | 0.953 | .93 | 63.9 | 0.944 | .93 | 62.7 | 0.954 | .94 | 64.8 |
| | 1.037 | .73 | 59.5 | 1.037 | .94 | -64.6 | 1.037 | .96 | -63.4 |
| | 1.037 | .94 | -65.5 | 1.037 | .62 | 60.5 | | | |

**Table 2. Example Modal Decomposition**

Figures 3b and 3c, which contain measurement noise, reveal similar mode behaviour to that in Figure 3a. While there is some difficulty in correctly modelling the spectrum when BT=50, the results for BT=100 are very similar to those in Figure 3a. Since in practice, BT will likely significantly exceed BT=100 for broadband detection, it appears that measurement noise should not be a major factor in limiting the mode identification.

## VI. Summary

The paper has considered modal decomposition as a means of directly separating plane wave signals from the more diffuse noise background in the display of broadband beams from a line array receiver. While low order modes can be successfully identified with plane wave signals of relatively high power, detection of weaker signals using higher order modes leads to problems such as mode splitting for the stronger signals. Indirect identification using the mode parameters may still lead to accurate estimates of signal angle and power. Measurement noise does not appear to be a limiting factor in this application. Note that since the method is aimed at identification of the plane wave signals, it is in fact a detection device.

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13. ABSTRACT

Detection of broadband signals is difficult in the sonar application due to the uncertainty in the broadband noise power and hence the uncertainty in the threshold required for detection. However, if the receiver has sufficient directivity, then it is possible to exploit the spatial characteristics of the broadband field to isolate plane wave signals from the more diffuse noise background. This paper considers a parameterization of the broadband bearing estimate from a line array receiver in terms of complex modes (Prony method). Identification of the modes associated with the plane wave arrivals is examined as a means of directly separating the signals from the noise. Only broadband beampowers are assumed available, so the bearing estimates are contaminated by the array response. Issues of mode number and estimation error are addressed by way of example.

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KEY WORDS

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